A BROADBAND MICROWAVE SINGLE SIDEBAND MODULATOR USING FERRITES — A FEASIBILITY STUDY

GEORGE M. SHELDON

A BROADBAND MICROWAVE SINGLE SIDEBAND MODULATOR USING FERRITES A FEASIBILITY STUDY

* * * * * *

George M. Sheldon

A BROADBAND MICROWAVE SINGLE SIDEBAND MODULATOR USING FERRITES

A FEASIBILITY STUDY

by

George Morgan Sheldon

Lieutenant, United States Navy

Submitted in partial fulfillment
of the requirements
for the degree of
MASTER OF SCIENCE

IN

ENGINEERING ELECTRONICS

United States Naval Postgraduate School
Monterey, California

1 9 5 5

This work is accepted as fulfilling
the thesis requirements for the degree of

MASTER OF SCIENCE

IN

ENGINEERING ELECTRONICS

from the

United States Naval Postgraduate School

PREFACE

As electronic production techniques have developed, need has arisen for test equipment that will do the job rapidly and reliably. One device needed is a broadband, automatic impedance measuring device for use at microwave frequencies. The input frequency to such a device could be swept over a waveguide band in order to obtain a plot of the impedance of a waveguide component versus frequency as rapidly as the signal source can be swept and the components to be measured interchanged.

This paper is concerned with the feasibility of one component of such a device; i.e., a stable, wideband single sideband modulator which can be used to shift the swept carrier frequency by an amount in the middle audio frequencies. Due to the small frequency shift desired relative to the carrier frequency involved, and the difficulties involved in single sideband modulation of a swept carrier frequency, normal methods of modulation and carrier and sideband suppression are not practical. Therefore, in order to obtain this component it is necessary to extend the work of John Cacheris [2] who has developed a successful single sideband modulator utilizing ferrites for use at a single X-band frequency.

In recent years, ferrites have been gaining a prominent place in microwave electronics due to their characteristics

when subjected to a D.C. magnetic field. Out of investigations into these characteristics have evolved such items as one way transmission systems, modulators and phase shifters.

The writer wishes to extend his grateful appreciation to the Hewlett-Packard Company, Frank Barnett of Hewlett-Packard, and Professor Carl Menneken of the U. S. Naval Postgraduate School for their invaluable assistance, encouragement, and cooperation in the research for this paper and in its preparation.

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

TABLE OF SYMBOLS AND ABBREVIATIONS

(Listed in the order of their use in the text)

db	- Decibel
f_c	- Carrier Frequency
f_m	- Modulation Frequency
SSM	- Single Sideband Modulator
DET	- Detector
E vector	- The Electric Field Vector of an Electromagnetic Wave
$\Delta\ 90^{\circ}$	- 90° Differential Phase Shift
$\Delta\ 180^{\circ}$	- 180° Differential Phase Shift
ω	- Angular Frequency
E_f	- Component of the Electric Field at the Carrier Frequency
E_{f+f_m}	- Component of the Electric Field at the Upper Sideband Frequency
E_{f-f_m}	- Component of the Electric Field at the Lower Sideband Frequency
ϵ	- Deviation of a Quarter Wave Plate From 90° Differential Phase Shift
δ	- Deviation of a Half Wave Plate From 180° Differential Phase Shift
α	- Differential Attenuation of a Half Wave Plate in Nepers
H vector	- The Magnetic Field Vector of an Electromagnetic Wave
n	- Index of Refraction
ϵ	- Dielectric Constant
μ	- Magnetic Permeability

γ - Gyromagnetic Ratio of the Electron

H_e - Effective Internal Magnetic Field

B_e - Effective Flux Density ($B_e = H_e \quad 4 M$)

M - Magnetization of the Medium

c - Velocity of Light

$\Delta \phi$. - Differential Phase Shift

B_s - Saturation Flux Density

B_r - Residual Flux Density

H_c - Coercive Force

P_c - Copper Loss

ρ - Volume Resistivity

P_h - Hysteresis Loss

w - Hysteresis Loss per Cycle

P_e - Eddy Current Loss

P_t - Total Losses

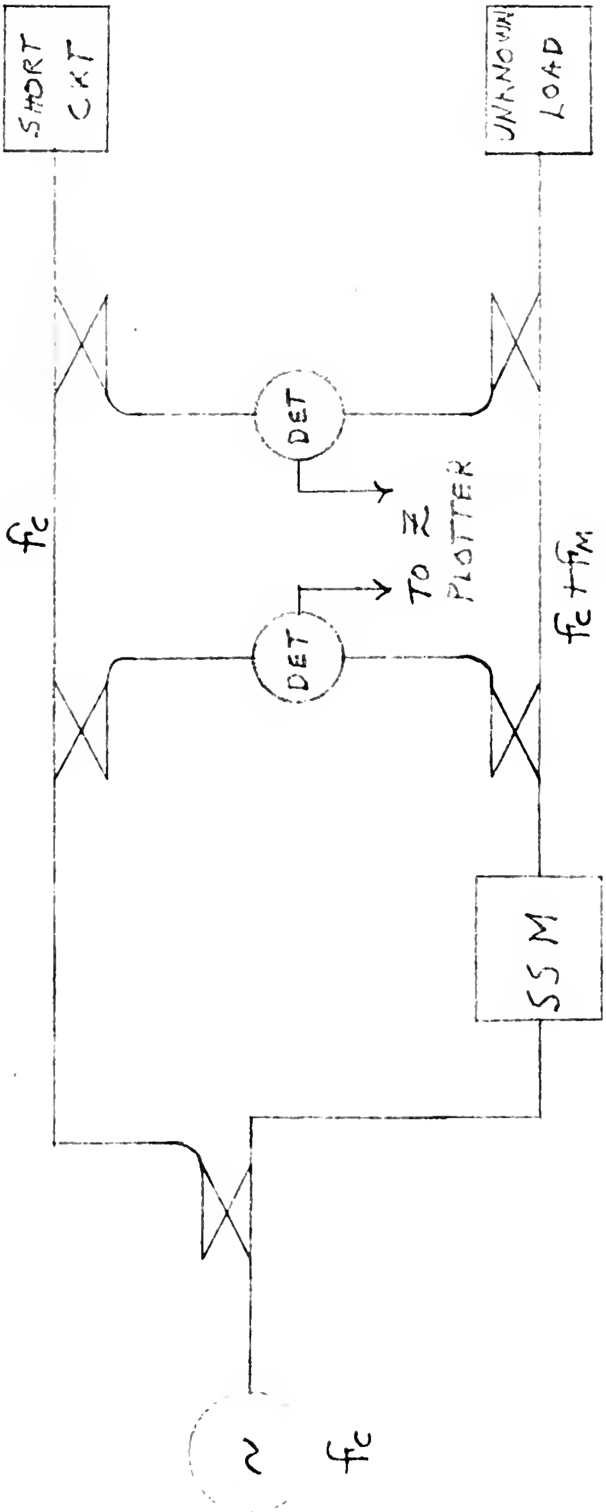
CHAPTER I

INTRODUCTION

1. An Automatic Impedance Measuring Device.

The basic concepts of an automatic impedance measuring device are illustrated in figure 1. The signal source would be a small klystron whose frequency could be swept over the entire waveguide band. The signal from the source is split in a directional coupler. A portion of this signal is sent down a waveguide to a known termination whose impedance is constant with frequency, in this case a short circuit. The other portion of the signal is sent through a single side-band modulator to an unknown load. The single sideband modulator is used to give a constant difference frequency between the two branches of the device while the microwave signal source is being swept in frequency.

The energy incident on the two loads is then sampled and combined in a detector whose output is the difference frequency. This output is to provide a comparison standard. The energy reflected from the two loads is also sampled and combined in another detector. The unknown load will control the phase and magnitude of the output of the second detector. This output when compared with the output of the first detector will provide information as to the phase and magnitude of the unknown impedance.



BASIC CONCEPT OF AUTOMATIC IMPEDANCE MEASURING DEVICE

FIGURE 1

The problem with which this paper is concerned is that of determining the feasibility of producing the microwave single sideband modulator. The requirements on the single sideband modulator as set forth by the Hewlett-Packard Company are that the carrier frequency will be suppressed by 20 to 30 db and that the lower sideband will be suppressed by at least 50 db. It is desired that the modulator meet these requirements over as much of the X-band (8.2 to 12.4 kilomegacycles per second) as is possible, and especially that these requirements be met over the lower half of the band which is the portion most used. It will be shown that this modulator can be constructed utilizing ferrites to cover the entire X-band.

2. Brief History of Previous Developments and Applications of Ferrites.

Polder [6] has analyzed the behavior of a saturated ferromagnetic material magnetized in an arbitrary direction with respect to the direction of propogation of an electromagnetic wave. This analysis was done both by use of macroscopic theory and Maxwell's equations and by quantum theory; however, both magnetic and dielectric losses were neglected in the development which therefore presents an idealized picture. This work has provided a basis for almost all further analytical work in the field.

Hogan [4] has described the development of a microwave "gyrator" utilizing the ferromagnetic Faraday effect in ferrites magnetized in the direction of propogation. He

has also extended Polder's theory to include both magnetic and dielectric losses for this direction of magnetization. Use of the Faraday effect has lead to the development of isolators and double sideband modulators which are now available commercially.

Lax, Button, and Roth [5] have described the utilization of ferrites in a rectangular waveguide magnetized perpendicularly to the direction of propogation to produce a variable phase shifter. The peculiar electric field distributions in the waveguide as described by Lax, Button, and Roth suggest the possibility of producing an isolator utilizing transverse magnetization. This idea is presently being pursued by engineers at several Laboratories.

Cacheris [2] has described the development of a microwave single sideband modulator using ferrites in a transverse magnetic field. The present project is an attempt to extend this work.

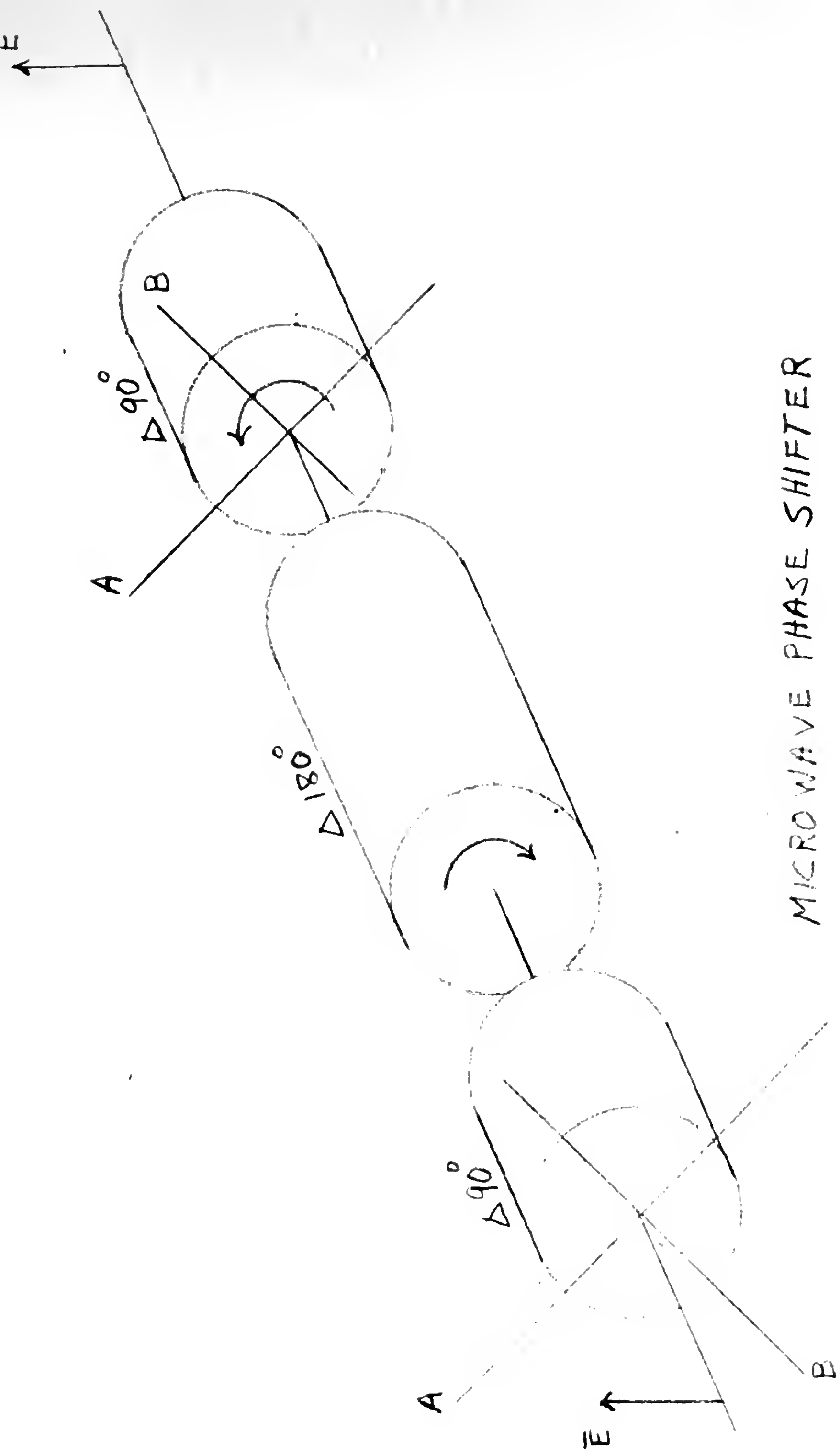
Suhl and Walker [7, 8, & 9] have expanded the theory given by Polder [6] and introduced the use of perturbation theory to describe the results obtained when the ferrite sample is small with respect to a wavelength or small with respect to the waveguide.

CHAPTER II

THEORY OF MICROWAVE PHASE SHIFTERS AND SINGLE SIDEBAND MODULATORS

1. A Mechanical Waveguide Phase Shifter and Single Sideband Modulator.

Fox [3] describes the development of a mechanical waveguide phase changer. This phase changer consists of two 90° differential phase shift sections and an 180° differential phase shift section as shown in figure 2. These sections correspond respectively to quarter wave and half wave plates in optics, which terminology will be used here. The first quarter wave plate converts the linearly polarized incident wave into a circularly polarized wave by retarding the component of the E vector in the direction of the B axis by 90° with respect to the component in the direction of the A axis. The half wave plate will reverse the sense of rotation of the circularly polarized wave back to a linearly polarized wave. The instantaneous angle of the output vector from the half wave plate will depend on the instantaneous angle of the input vector with respect to the principal axes of the section. Thus by rotating the half wave section, the instantaneous output polarization can be rotated. As this polarization is rotated, the phase of the output of the second quarter wave plate will vary. By rotating the half wave plate one half revolution, the output phase from the phase changer



$\bar{E}'$
B
A
$\Delta 90^\circ$
$\Delta 180^\circ$
$\Delta 90^\circ$
A
B
$\bar{E}$
MICRO WAVE PHASE SHIFTER

FIGURE 2

is varied through 360°. If the half wave plate is rotated at a constant angular velocity, it will cause a fixed increase or decrease in frequency of the output from that of the input. This frequency change is equal to twice the frequency of rotation of the half wave plate. The modulating frequency in this case is limited by the permissible speed of rotation of the half wave plate.

2. Analysis of a Waveguide Phase Shifter or Single Sideband Modulator.

The following analysis is made assuming the use of lossless elements and exact 90° and 180° differential phase shift sections.

Assume a linearly polarized wave incident on the first quarter wave plate whose plane of polarization makes an angle of 45° with both axes A and B as indicated by the vector E in figure 2. This wave will have components in the direction of the A and B axes of:

$$\left.\begin{aligned} E_A &= \cos \omega t \\ E_B &= -\cos \omega t \end{aligned}\right\} \quad \text{normalized} \qquad (1)$$

The output of the quarter wave plate will be:

$$E_A = \cos \omega t \qquad (2)$$

$$E_B = -\cos(\omega t - 90°) = -\sin \omega t$$

The input to the half wave plate, where O is the angle of orientation of the principle axis X of the half wave plate with respect to the A direction, is:

$$E_x = E_A \cos\theta - E_B \sin\theta$$
$$E_y = -E_A \sin\theta - E_B \cos\theta$$
$$E_x = \cos\theta \cos\omega t + \sin\theta \sin\omega t$$
$$E_y = -\sin\theta \cos\omega t + \cos\theta \sin\omega t \qquad (3)$$
$$E_x = \cos(\omega t - \theta)$$
$$E_y = \sin(\omega t - \theta)$$

The output of the half wave plate is:

$$E_x = \cos(\omega t - \theta)$$
$$E_y = \sin(\omega t - \theta - 180^\circ) = -\sin(\omega t - \theta) \qquad (4)$$

The input to the second quarter wave plate is:

$$E'_A = E_x \cos\theta - E_y \sin\theta$$
$$E'_B = E_x \sin\theta + E_y \cos\theta$$

$$E'_A = \cos\theta \cos(\omega t - \theta) + \sin\theta \sin(\omega t - \theta)$$
$$E'_B = -\left[\sin\theta \cos(\omega t - \theta) + \cos\theta \sin(\omega t - \theta)\right] \qquad (5)$$
$$E'_A = \cos(\omega t - 2\theta)$$
$$E'_B = -\sin(\omega t - 2\theta)$$

The output of the second quarter wave plate is then:

$$E'_A = \cos(\omega t - 2\theta)$$
$$E'_B = -\sin(\omega t - 2\theta - 90^\circ) = \cos(\omega t - 2\theta) \qquad (6)$$

If the half wave plate is then rotated at a constant angular velocity of ω_h in the direction of minus θ, the output of the second quarter wave plate becomes

$$E'_A = \cos(\omega + 2\omega_h)t$$
$$E'_B = \cos(\omega + 2\omega_h)t \qquad (7)$$

which will add to give a linearly polarized output wave of the form:

$$E' = \sqrt{2}\cos(\omega + 2\omega_m)t. \tag{8}$$

It can be seen that this is an equation for single sideband modulation where the carrier and lower sideband frequencies are completely suppressed.

3. The Effect of Errors in a Microwave Single Sideband Modulator.

Barnett [1] has shown that the errors in a microwave single sideband modulator are due to deviations in the quarter wave plates and the half wave plate from the ideal. Deviations in the half wave plate contribute only to the appearance of the carrier frequency in the output of the modulator. Deviations in the quarter wave plates contribute to the appearance of both the carrier and lower sideband frequencies in the output of the modulator. The following equations will show the effect of these errors:

$$\frac{E_f}{E_{f+f_m}} = 2\tan\frac{1}{2}\epsilon\sqrt{\frac{\sin^2\frac{1}{2}\delta + \sinh^2\frac{1}{2}\alpha}{\cos^2\frac{1}{2}\delta + \sinh^2\frac{1}{2}\alpha}} \tag{9}$$

$$\frac{E_{f-f_m}}{E_{f+f_m}} = \tan^2\frac{1}{2}\epsilon \tag{10}$$

These equations can be derived by making the differential phase shift of the quarter wave plates $90^\circ \pm \epsilon$ and by making the differential phase shift of the half wave plate $180^\circ \pm \delta$ and by inserting a differential attenuation of $e^{-\alpha}$ in the half wave plate in the foregoing analysis of the waveguide phase

shifter or single sideband modulator.

4. A Ferrite Differential Phase Shift Section.

Since there are limitations on the speed of rotation of the half wave plate in using a mechanical wave guide phase shifter to make a single sideband modulator, it is proposed to make a half wave plate using a piece of ferrite in a circular waveguide in a rotating transverse magnetic field. The rotating transverse magnetic field will cause rotation of the half wave plate as can be seen from the following theory. The magnetic field can be made to rotate at the desired angular velocity by driving a four pole magnet with a two phase voltage of the desired frequency.

The theory of the ferrite differential phase shift section as given by Cacheris [2] is as follows:

> An infinitely large ferromagnetic medium which is homogeneously magnetized in the Y direction becomes doubly refracting when plane electromagnetic waves are propogated in the Z direction. For two linearly polarized waves with H vectors respectively parallel and perpendicular to the applied D.C. magnetic field, the indices of refraction are given by:
>
> $$n_{\parallel}^2 = \epsilon \tag{11}$$
>
> $$n_{\perp}^2 = \frac{\epsilon(\mu^2-\alpha^2)}{\mu} \tag{12}$$
>
> If ferromagnetic losses and crystal anisotropy fields are ignored, μ and α are given by:
>
> $$\mu = \frac{\gamma^2 He\, Be - \omega^2}{\gamma^2 He^2 - \omega^2} \tag{13}$$
>
> and
>
> $$\alpha = \frac{4\pi M \omega \gamma}{\gamma^2 He^2 - \omega^2} \tag{14}$$
>
> Then
>
> $$n_{\parallel}^2 = \epsilon = \epsilon \mu_{\parallel} \tag{15}$$

and

$$n_\perp^2 = \epsilon\left[\frac{\gamma^2 B_e^2 - \omega^2}{\gamma^2 H_e B_e - \omega^2}\right] = \epsilon\mu_\perp \tag{16}$$

The effective permeability of the wave whose H vector is parallel to the D.C. magnetic field, H_a, is unity, since $n^2 = \mu\epsilon$. It is independent of the magnitude of H_a. The effective permeability of the wave whose H vector is perpendicular to the D.C. field is given by:

$$\mu_\perp = \left[\frac{\gamma^2 B_e^2 - \omega^2}{\gamma^2 H_e B_e - \omega^2}\right] \tag{17}$$

The perpendicular permeability becomes infinite when $\omega^2 = \gamma^2 H_e B_e$, not when $\omega^2 = \gamma^2 H_e^2$ as for the medium magnetized in the direction of propogation. This agrees with the theory.

Let $\mu_\perp = 1 + \Delta\mu_\perp$

$$\Delta\mu_\perp = \left[\frac{\gamma^2 B_e - \omega^2}{\gamma^2 H_e B_e - \omega^2} - 1\right] \tag{18}$$

$$\Delta\mu_\perp = \frac{\gamma^2 B_e^2 - \gamma^2 H_e B_e}{\gamma^2 H_e B_e - \omega^2} \tag{19}$$

If $\gamma^2 H_e B_e$ is much smaller than ω^2, the above equation can be simplified to

$$\Delta\mu_\perp = \frac{-\gamma^2 B_e^2 + \gamma^2 H_e B_e}{\omega^2} \tag{20}$$

and

$$\mu_\perp = \left[1 + \frac{\gamma^2 H_e B_e - \gamma^2 B_e^2}{\omega^2}\right] \tag{21}$$

The differential phase shift of the wave with its magnetic vector parallel to H_a, with respect to the wave whose magnetic vector is perpendicular to H_a, is given by

$$\Delta\Phi = \omega l \frac{(\epsilon_r)^{1/2}}{c}\left[\mu_{\parallel}^{1/2} - \mu_\perp^{1/2}\right] \tag{22}$$

or

$$\Delta\Phi \cong l\frac{(\epsilon_r)^{1/2}}{2c}\left[\frac{\gamma^2 B_e^2 - \gamma^2 H_e B_e}{\omega}\right] \tag{23}$$

Although the theory has been derived for a saturated infinite medium with no losses, it appears that the analysis approximates the waveguide case since double refraction has been observed experimentally.

Cacheris [2] experimentally verified the foregoing theory and successfully built a single sideband modulator to shift a microwave carrier frequency of 9.375 kilomegacycles by plus or minus 20 kilocycles.

CHAPTER III

EXPERIMENTAL PROCEDURES AND RESULTS

1. Extent of Experiments.

Due to the amount of time available for this study, it was found that the effects of operating the ferrite half wave plate in a rotating magnetic field could not be investigated. However, the principle active portion of the broadband single sideband modulator is the half wave plate. Experiments to determine the feasibility of this component utilizing D.C. magnetic fields were completed and are described below. It is believed that the experiments conducted will adequately determine the feasibility of the modulator.

2. Differential Phase Shift Measurements.

The purpose, therefore, of the measurements was to find a ferrite sample and mounting for that sample that would provide a minimum deviation from a 180° differential phase shift for a constant D.C. magnetizing current while the test frequency is varied throughout the X-band (8.2 to 12.4 kilomegacycles per second). To obtain the required data, differential phase shift and insertion loss measurements were made in a phase null bridge as shown in figure 3. The standing wave ratio due to the mismatch of the ferrite to the waveguide was measured by inserting a slotted waveguide section ahead of the first transition section in the bridge.

The D.C. magnetic field was first applied across the ferrite in a direction parallel to the E vector of the incident electromagnetic wave and then perpendicular to the E vector. The magnetic field was varied from the value of residual magnetization of the magnet yoke to a value required to give over 180° of differential phase shift. These tests were then repeated at frequencies throughout the X-band for each ferrite sample.

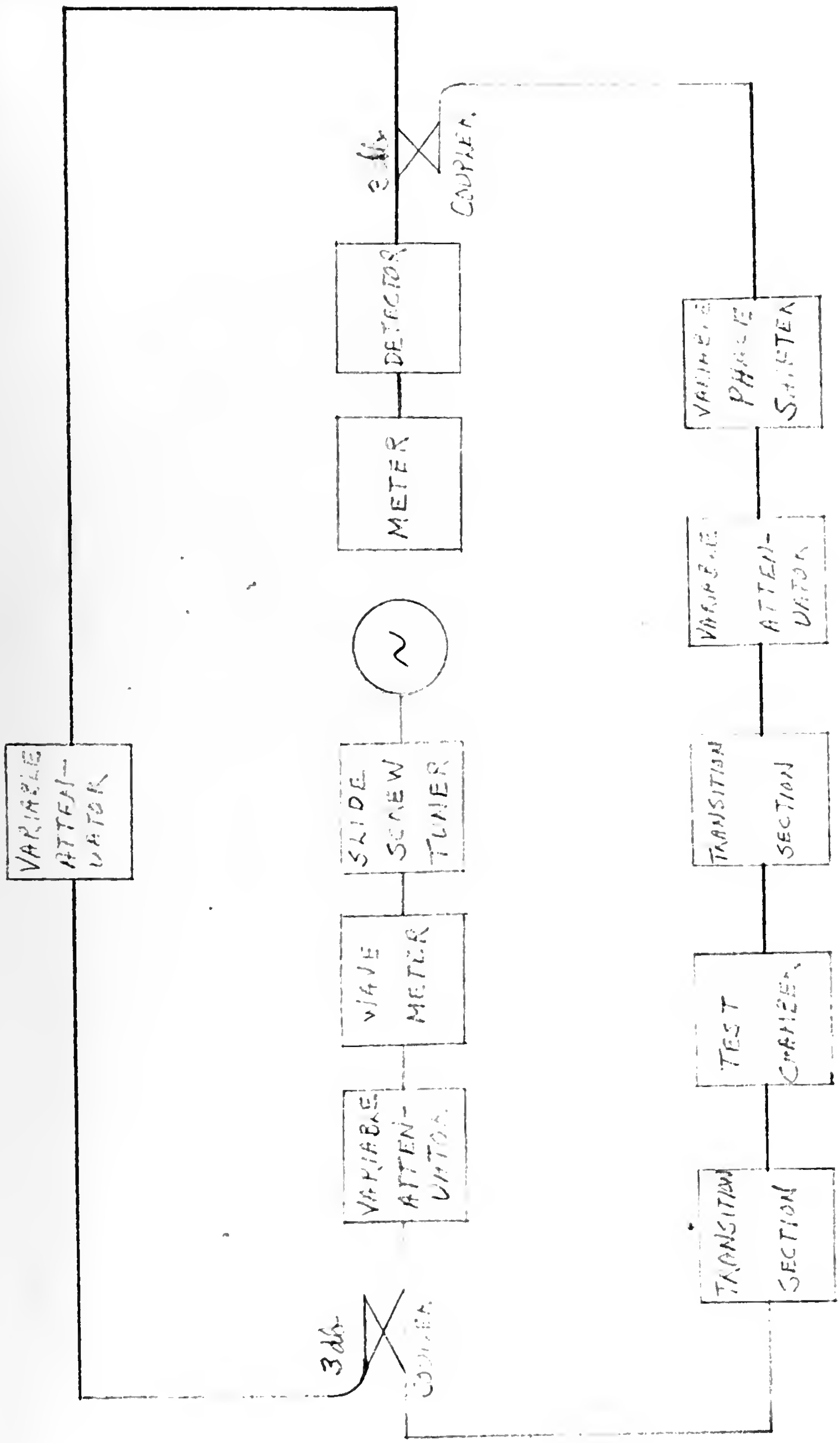
3. Ferrite Samples.

The ferrite samples used were Ferramic MF-1331 (now available commercially as Ferramic "R1") made by General Ceramics Corporation of Keasbey, New Jersey. Ferramic MF-1331 was developed specifically for microwave applications.

Electrical characteristics and sample sizes are listed in tables 1 and 2, and the various methods of mounting the ferrites in the waveguide are shown in figures 4, 5, and 6.

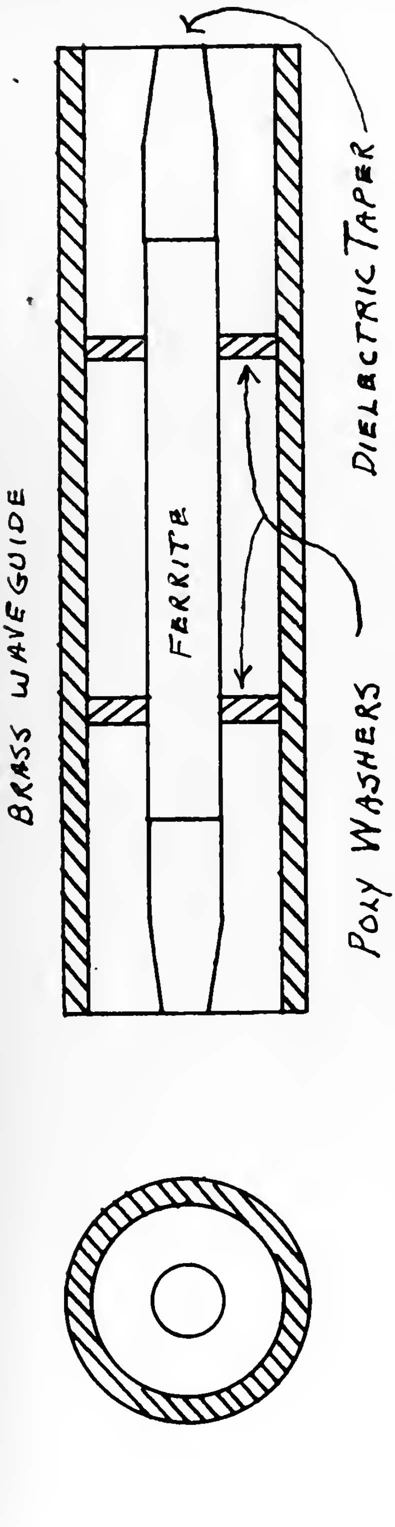
4. Line of Investigation.

It was felt that the most promise in obtaining a broad band device was to investigate various mountings for the ferrite samples including methods of concentrating the magnetic flux at the ferrite and minimizing air gaps in the magnetic structure. The mounting shown in figure 4 was used primarily to become familiar with the techniques of measurement and the general behaviour of the ferrites in a transverse field. The other mountings allowed the magnet pole pieces to be close to the ferrite and thus minimized the air



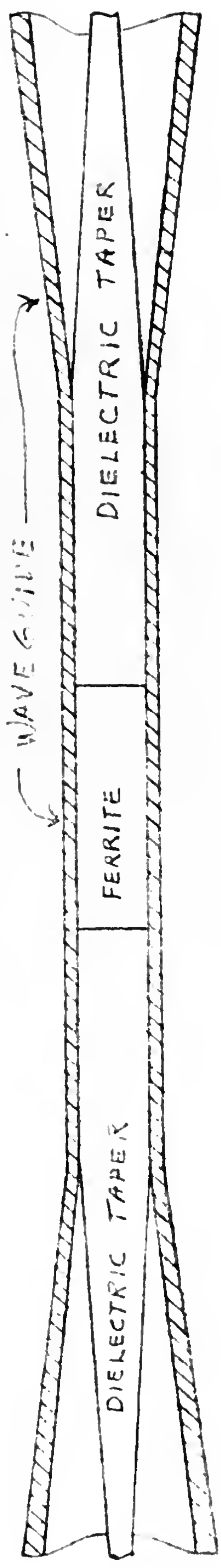
NULL BRIDGE FOR MEASUREMENT OF PHASE SHIFT AND INSERTION LOSS

FIGURE 3



MOUNTING FOR FERRITE SAMPLE 1

FIGURE 4

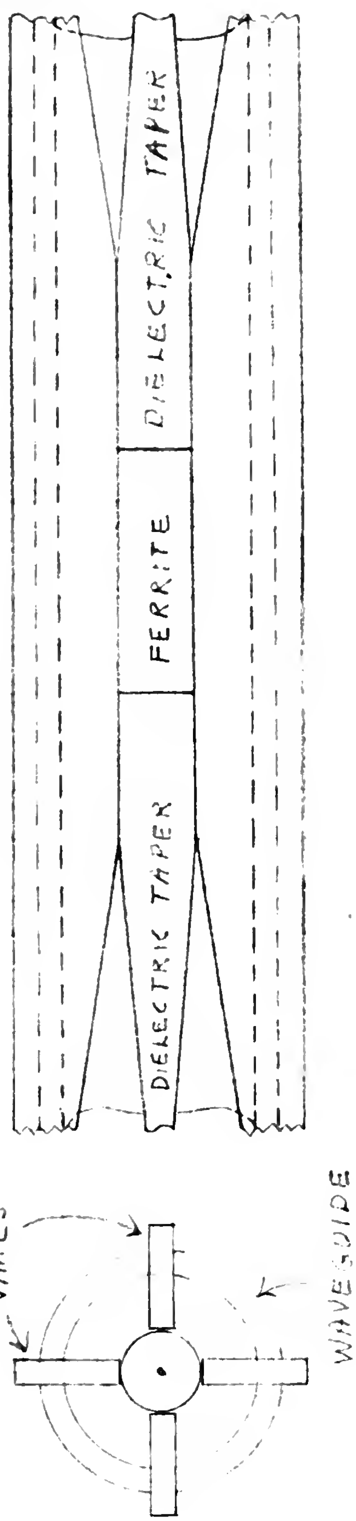


SECOND MOUNTING FOR FERRITE SAMPLE 2

FIRST MOUNTING SAME AS SHOWN

IN FIGURE 4.

FIGURE 5



MOUNTING FOR FERRITE SAMPLES THREE, FOUR, AND FIVE

FIGURE 6

gap which permitted use of smaller magnetizing currents in the electromagnet. The iron vane mountings shown in figure 6 are believed to be the most promising, since they concentrate the magnetic field at the ferrite, and also tend to concentrate the R.F. fields in the ferrite.

5. Analysis of Experimental Data.

As can be seen from the graphs of field current required for a 180° differential phase shift plotted against frequency, there are wide variations in the field currents required for the 180° differential phase shift. Also, it will be noted that the average magnitude of the field current varies considerably from sample to sample. This is due primarily to differences in the reluctance of the magnetic path from mounting to mounting.

Sample two, mounted in a completely filled waveguide as shown in figure 5, followed the inverse frequency characteristics as predicted for an infinite medium by equation (23). The other samples mounted in the iron vane mountings as shown in figure 6 did not follow this prediction over the frequency range tested. In fact, samples 4 and 5 showed, in general, the opposite characteristic. While not verified analytically, it is felt that this corresponds to perterbation effects as the sample diameter is made small with respect to a wavelength.

Mountings 1 and 2 for sample 4 vary only in the air gap introduced into the magnetic circuit. Mounting 1 has an air gap of about 0.074 inches between the ferrite and the iron

vanes; while mounting 2 has the iron vanes projecting into the ferrite sample.

Of all the samples tested, sample five showed the flattest differential phase shift response with changes in frequency. Therefore, it was felt that this sample would give the best chance of obtaining a broadband single sideband modulator.

The deviation in the available quarter wave plates is six degrees or less. These quarter wave plates are the ones that are used in the Hewlett-Packard mechanical waveguide phase shifter. The maximum deviations from a lossless half wave plate for sample five are 41^{o} in differential phase shift and 3.6 db or 0.415 nepers in differential attenuation. Both of these maximum deviations occurred at 12.4 kilomegacycles. Substituting this data in equations (9) and (10) one obtains:

$$\frac{E_f}{E_{f+f_m}} = 2\tan 3^\circ \sqrt{\frac{\sin^2 20.5^\circ + \sinh^2 0.202}{\cos^2 20.5^\circ + \sinh^2 0.202}} = 0.044$$

and

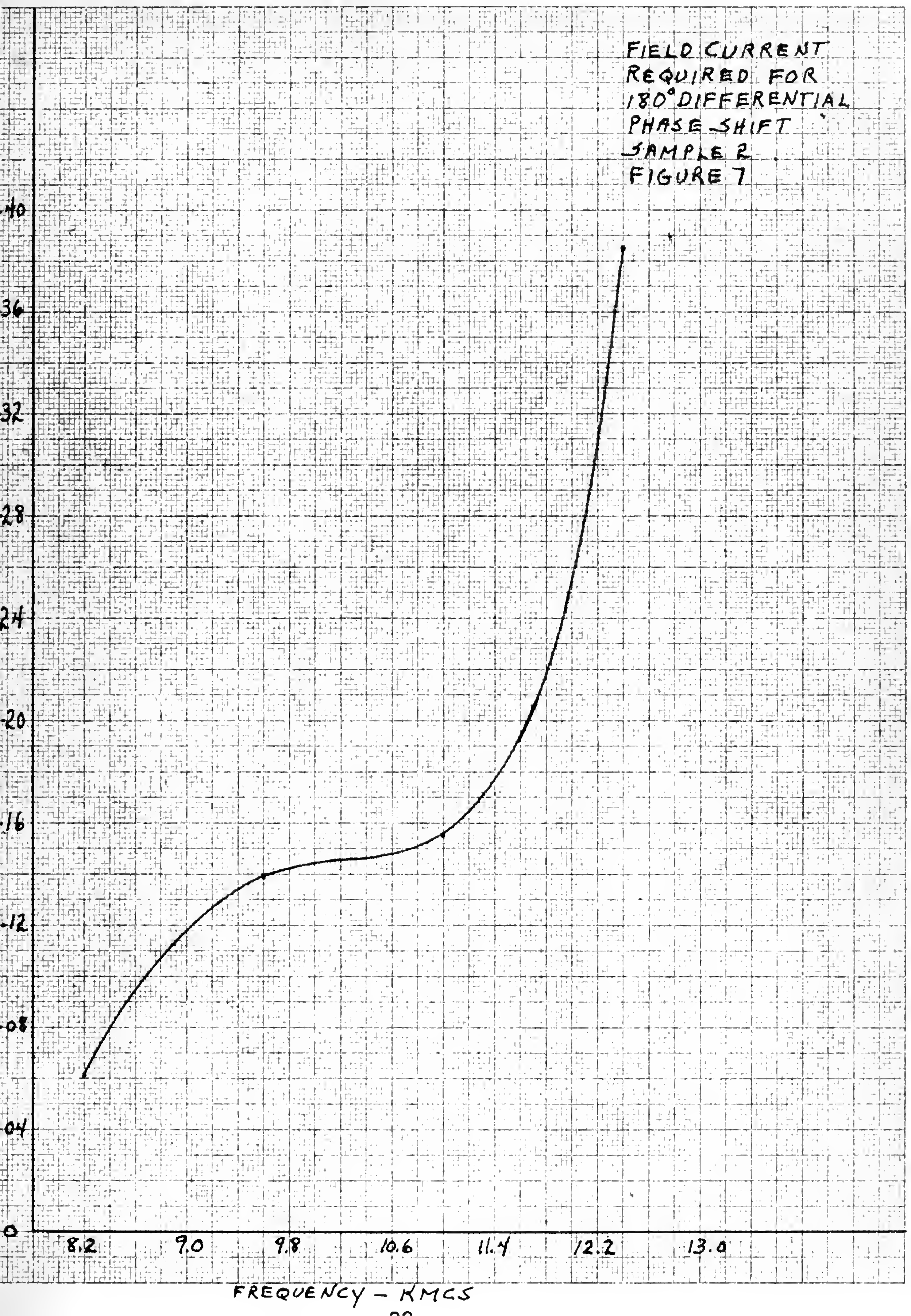
$$\frac{E_{f-f_m}}{E_{f+f_m}} = \tan^2 3^\circ = 0.00274$$

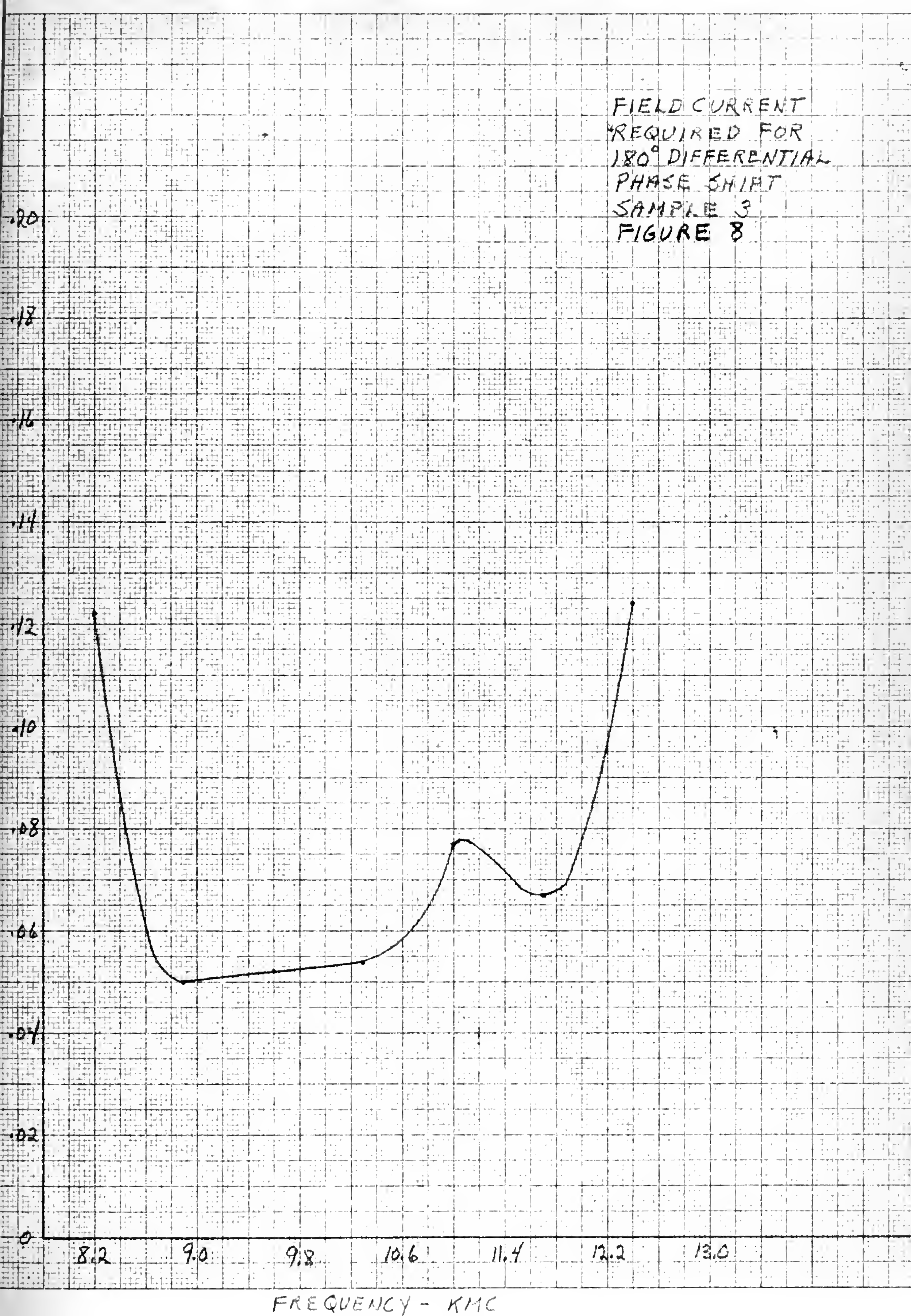
This corresponds to a carrier suppression of 27.7 db and a lower sideband suppression of 51.3 db. This exceeds the maximum requirements of 20 db carrier suppression and 50 db lower, sideband suppression for the automatic impedance measuring device.

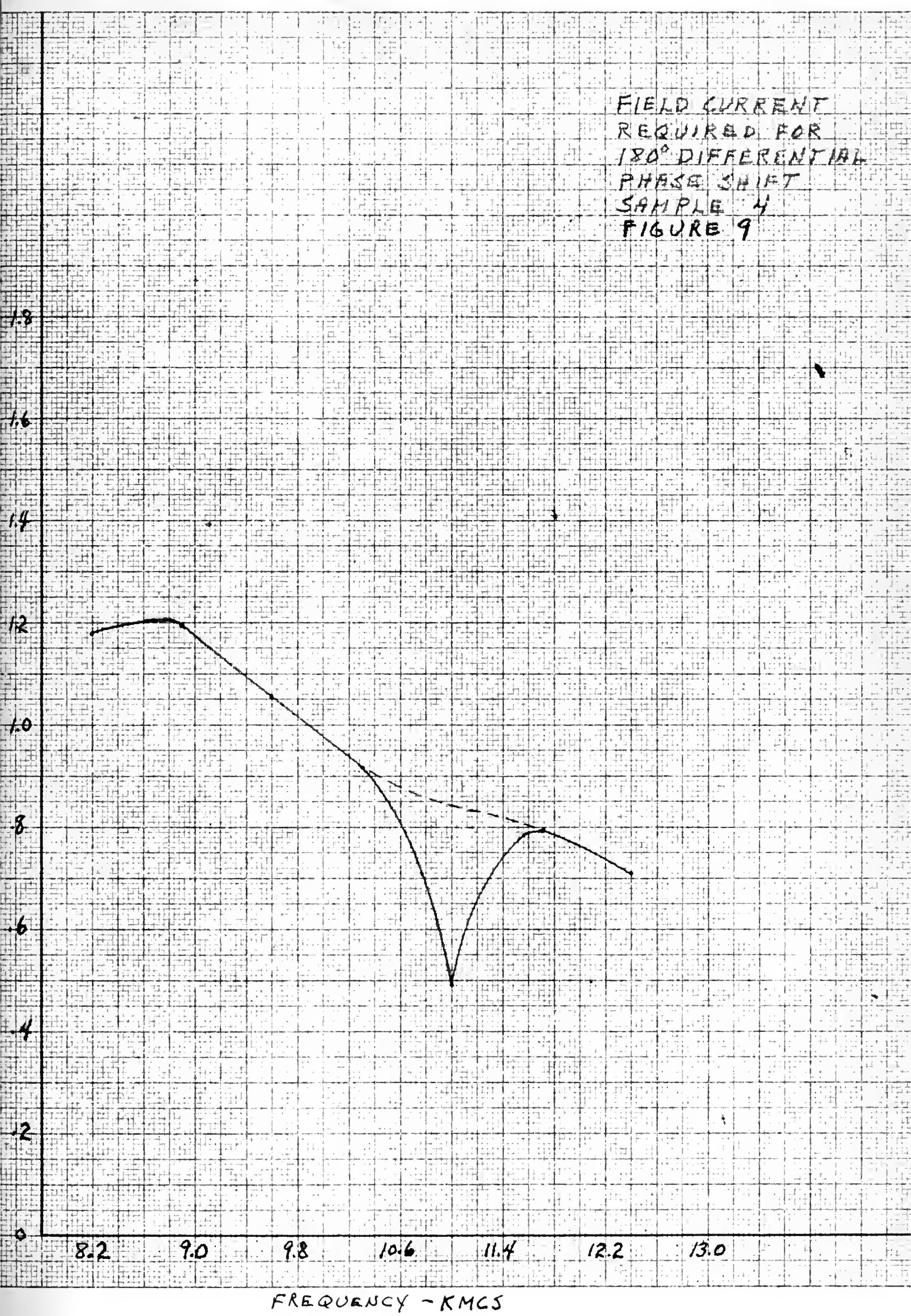
All samples tested showed a fairly broad variation of insertion loss over the waveguide band for 180^{o} of differential

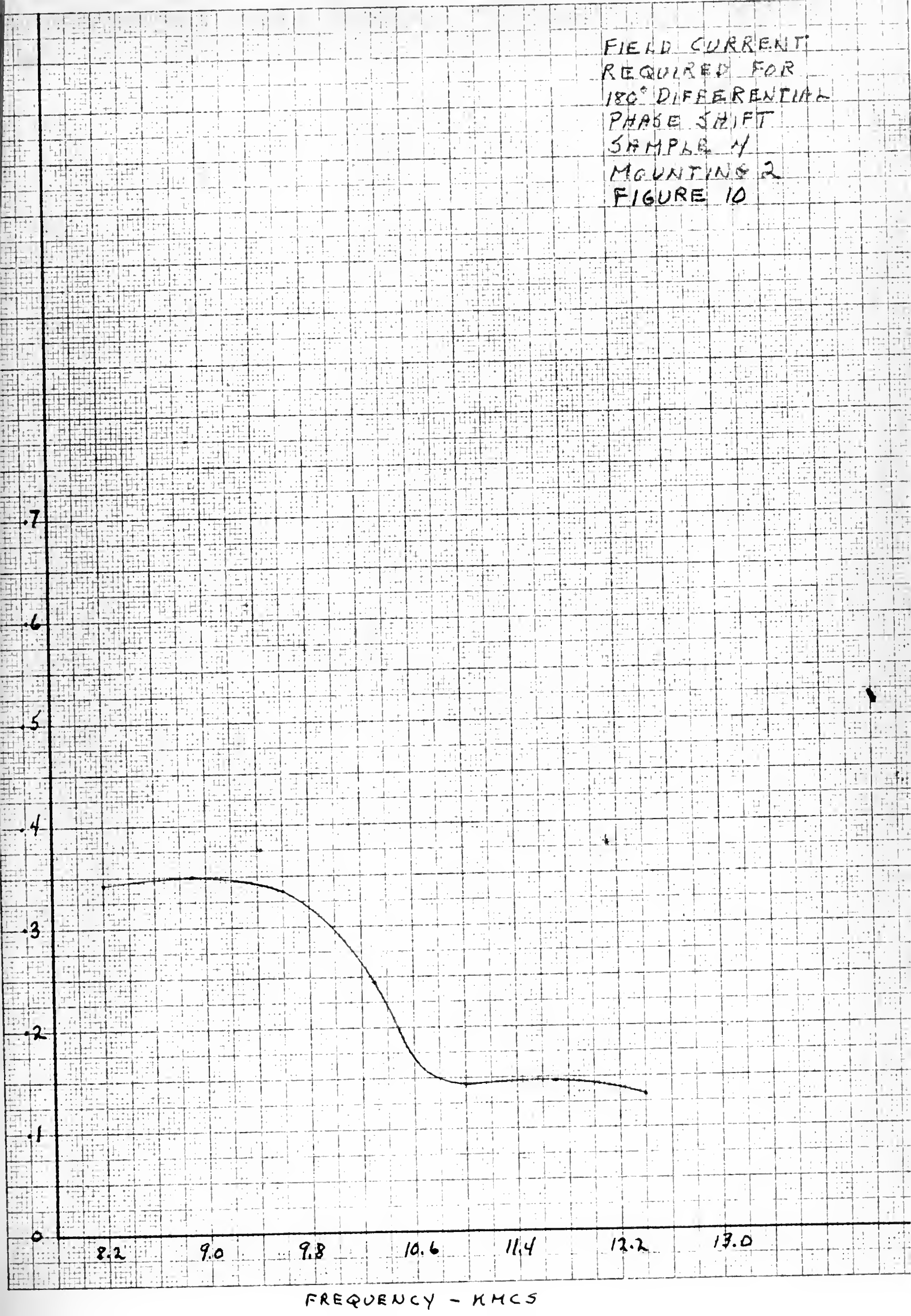
phase shift. The samples also showed wide variations at single frequencies as the D.C. magnetic field was varied. These losses and variations can be explained by changes in the nearness to ferromagnetic resonance when the D.C. magnetic field is parallel to the R.F. E vector, and by spurious modes propogating in the high dielectric tapers and in the ferrite.

Sample five showed high standing wave ratios of the order of 1.5 to 2.0 over most of the measured fields and frequencies. However this sample was mounted without the use of tapered matching sections. Use of tapers with a dielectric constant of about 10 would materially improve these standing wave ratios. Judging from previous experience, standing wave ratios of 1.1 to 1.3 should be achieved over the entire X-band using the matching tapers.


FIELD CURRENT
REQUIRED FOR
180° DIFFERENTIAL
PHASE SHIFT
SAMPLE 2
FIGURE 7
.40
.36
.32
.28
.24
.20
.16
.12
.08
.04
0
8.2
9.0
9.8
10.6
11.4
12.2
13.0
FREQUENCY – KMCS

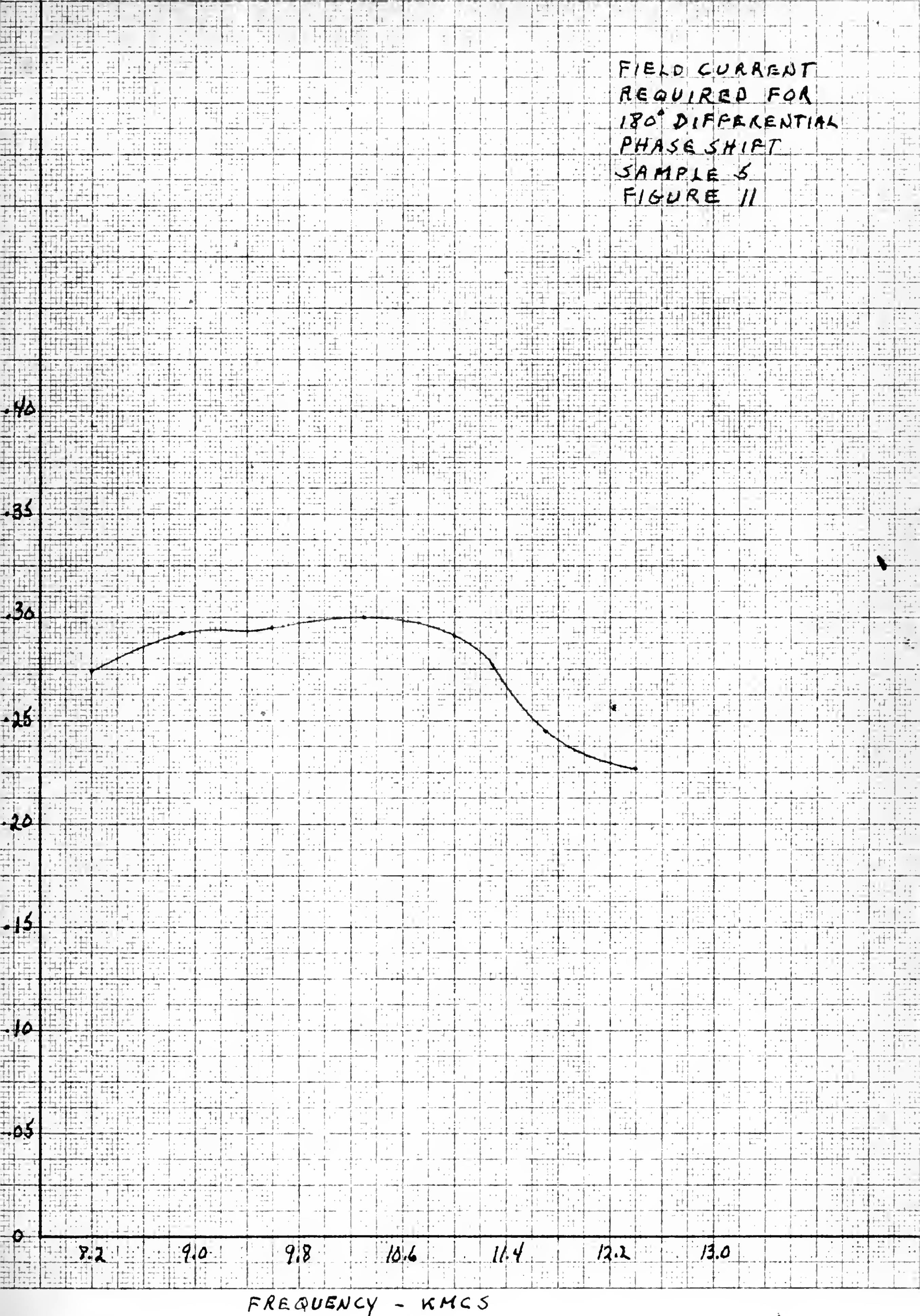

FIELD CURRENT REQUIRED FOR 180° DIFFERENTIAL PHASE SHIFT SAMPLE 3
FIGURE 8
.20
.18
.16
.14
.12
.10
.08
.06
.04
.02
0
8.2
9.0
9.8
10.6
11.4
12.2
13.0
FREQUENCY - KMC


FIELD CURRENT
REQUIRED FOR
180° DIFFERENTIAL
PHASE SHIFT
SAMPLE 4
FIGURE 9
1.8
1.6
1.4
1.2
1.0
.8
.6
.4
.2
0
8.2
9.0
9.8
10.6
11.4
12.2
13.0
FREQUENCY - KMCS

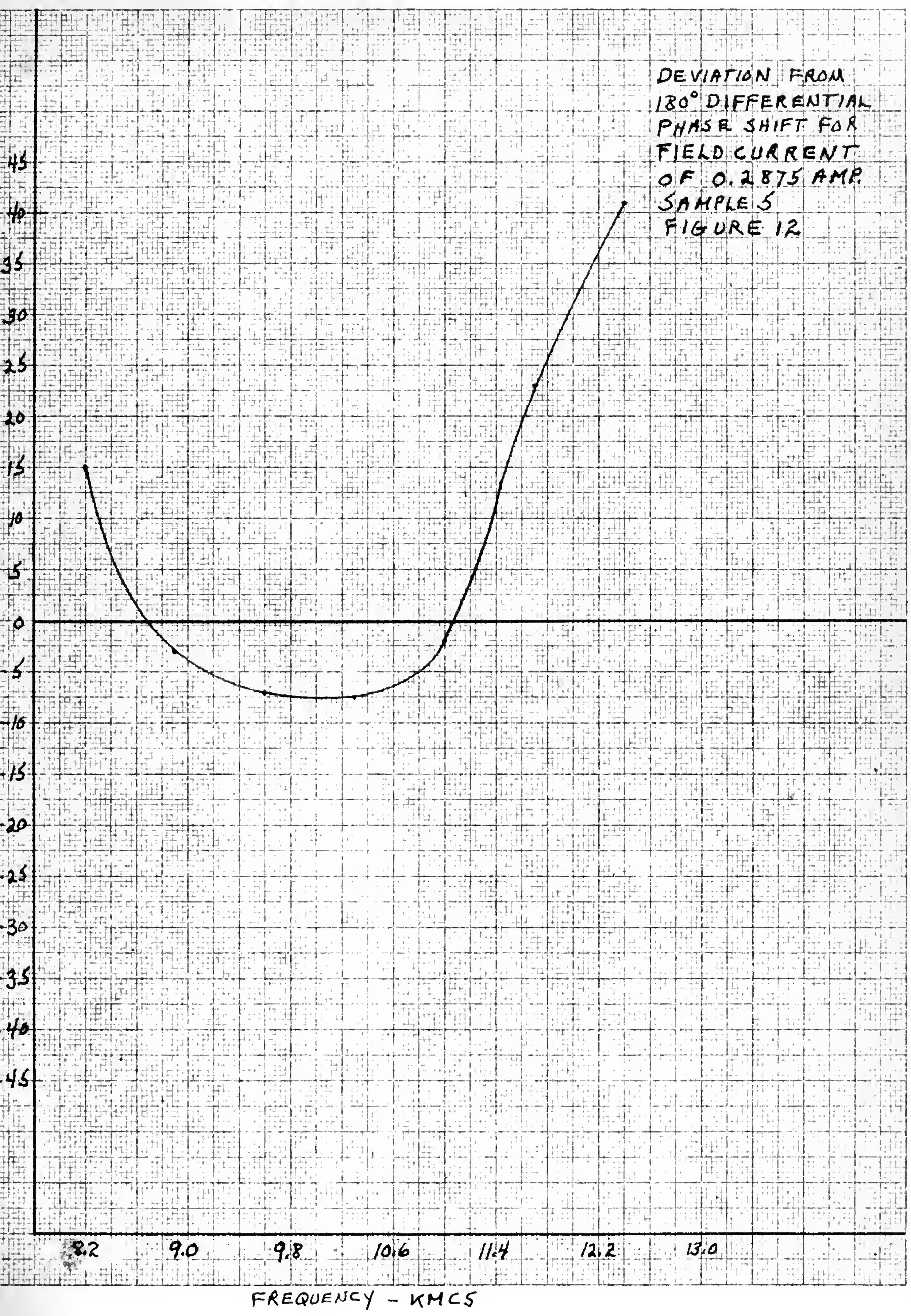


FIELD CURRENT
REQUIRED FOR
180° DIFFERENTIAL
PHASE SHIFT
SAMPLE 1
MOUNTING 2
.7
.6
.5
.4
.3
.2
.1
0
8.2
9.0
9.8
10.6
11.4
12.2
13.0
FREQUENCY - KMCS

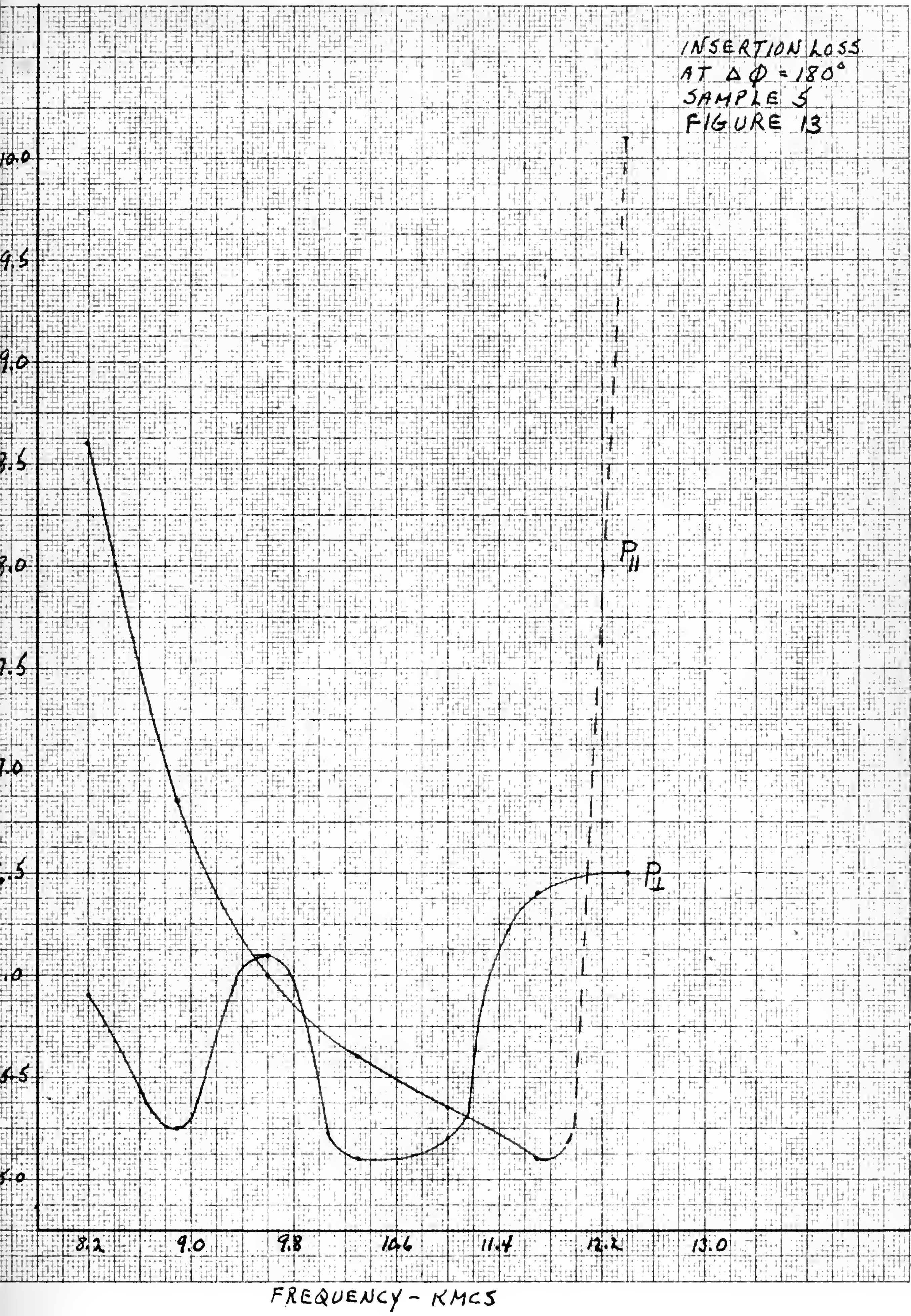
FIGURE 10



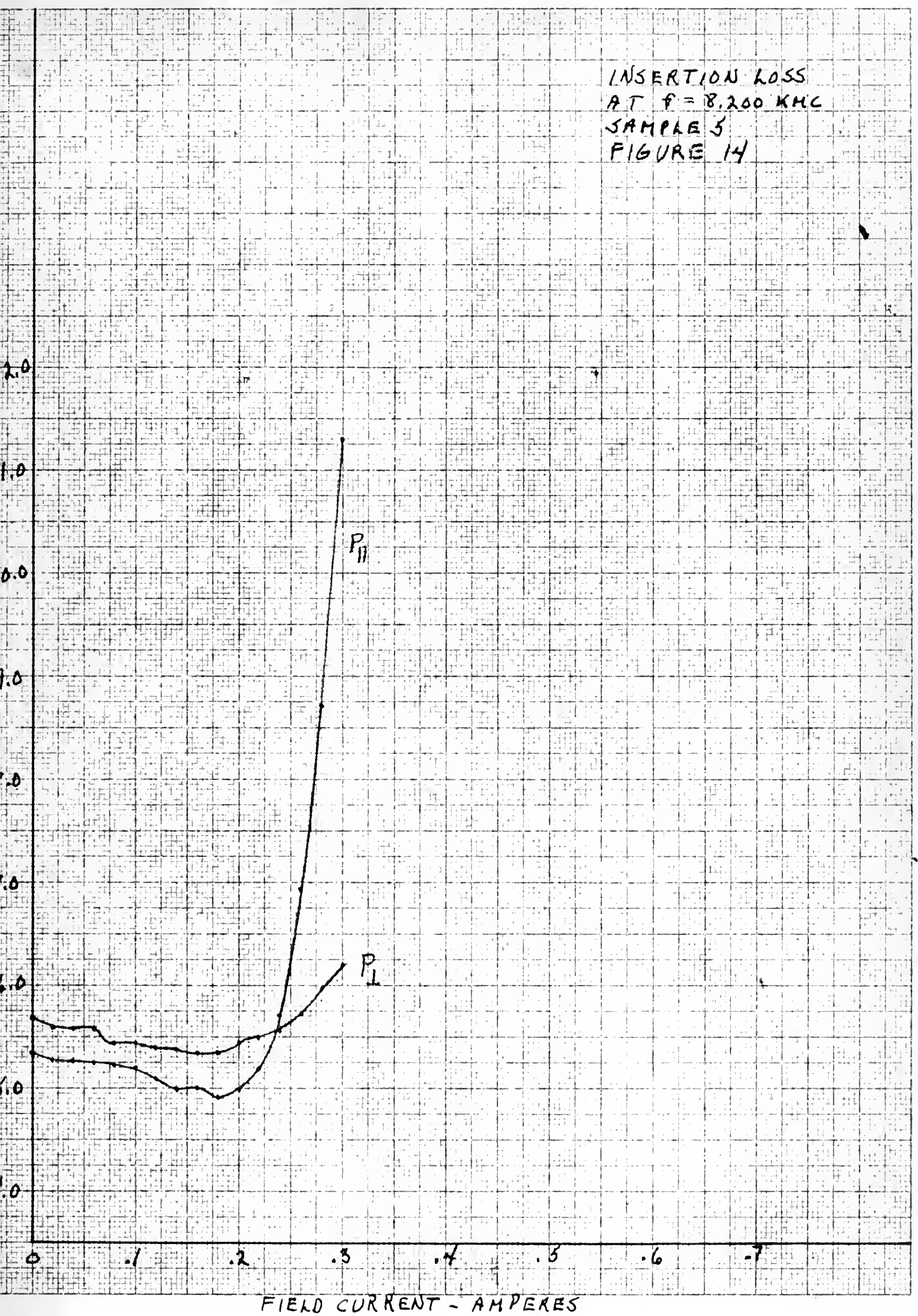
FIELD CURRENT
REQUIRED FOR
180° DIFFERENTIAL
PHASE SHIFT
SAMPLE 6
FIGURE 11
.40
.35
.30
.25
.20
.15
.10
.05
0
8.2
9.0
9.8
10.6
11.4
12.2
13.0
FREQUENCY - KMCS



DEVIATION FROM 180° DIFFERENTIAL PHASE SHIFT FOR FIELD CURRENT OF 0.2875 AMP.
SAMPLE 5
FIGURE 12
45
40
35
30
25
20
15
10
5
0
-5
-10
-15
-20
-25
-30
-35
-40
-45
8.2
9.0
9.8
10.6
11.4
12.2
13.0
FREQUENCY - KMCS



INSERTION LOSS
AT ΔΦ = 180°
SAMPLE 5
FIGURE 13
10.0
9.5
9.0
8.5
8.0
7.5
7.0
6.5
6.0
5.5
5.0
8.2
9.0
9.8
10.6
11.4
12.2
13.0
$P_{\parallel}$
$P_{\perp}$
FREQUENCY - KMCS


INSERTION LOSS
AT f = 8.200 KMC
SAMPLE 5
FIGURE 14
$P_{\parallel}$
$P_{\perp}$
0
.1
.2
.3
.4
.5
.6
.7
FIELD CURRENT - AMPERES

CHAPTER IV

CONCLUSIONS

1. Feasibility of Constructing a Broadband Single Sideband Modulator Using Ferrites.

The data obtained shows that such a broadband single sideband modulator is feasibile throughout the X-band for use in an automatic impedance measuring device requiring carrier suppression of at least 20 db and lower sideband suppression of at least 50 db. Samples three and four would each permit coverage of the entire band using constant power to the rotating magnetic field.

One problem not considered is the effect of hysteresis in the magnetic circuit introducing unwanted components in the output. It is believed that use of materials such as permalloy or supermalloy that have a very low coercive force and operating these materials below the knee of the saturation curve for the magnet yoke would minimize these effects so that they would cause no trouble. Care would have to be exercised in exciting the magnetic field so that its magnitude would remain constant during the rotation of the field. Considering the work of Cacheris [2], it is believed that this would be a minor problem.

The order of magnitude of power levels shown in Appendix II are readily obtainable, and therefore driving power requirements for the magnetic field are no problem. Also, since the

R.F. power used is measured in milliwatts, the R.F. power handling capabilities of the device are entirely adequate.

BIBLIOGRAPHY

1. Barnett, F. — Theory of Errors for a Microwave Single Sideband Modulator
   Hewlett-Packard Unpublished Memorandum, March 1955

2. Cacheris, J. — Microwave Single Sideband Modulator Using Ferrites
   Proc. I.R.E., Vol. 42, No. 8
   pp 1242-1247, August 1954

3. Fox, A. G. — Adjustable Waveguide Phase Changer
   Proc. I.R.E., Vol. 35, No. 12
   pp 1489-1498, December 1947

4. Hogan, C. L. — The Ferromagnetic Faraday Effect at Microwave Frequencies and Its Applications - The Microwave Gyrator
   B.S.T.J., Vol. 31, No. 1
   pp 1-31, January 1952

5. Lax, B.; Button, K. J.; and Roth, L. M. — Ferrite Phase Shifters in Rectangular Waveguide
   J. App. Phys., Vol. 25, No. 11
   pp 1413-1421, November 1954

6. Polder, H. — On the Theory of Ferromagnetic Resonance
   Phil. Mag., Vol. 40, No. 300
   pp 99-114, January 1949

7. Suhl, H. and Walker, L. R. — Topics in Guided-Wave Propogation Through Gyromagnetic Media Part I - The Completely Filled Cylindrical Guide
   B.S.T.J., Vol. 33, No. 3
   pp 579-659, May 1954

8. Suhl, H. and Walker, L. R. — Topics in Guided-Wave Propogation Through Gyromagnetic Media Part II - Transverse Magnetization
   B.S.T.J., Vol. 33, No. 4
   pp 939-986, July 1954

9. Suhl, H. and Walker, L. R. — Topics in Guided-Wave Propogation Through Gyromagnetic Media Part III - Perturbation Theory and Miscellaneous Results
   B.S.T.J., Vol. 33, No. 5
   pp 1133-1194, September 1954

APPENDIX I

TABLES

TABLE I

PRELIMINARY DATA SHEET

Ferramic MF-1331

ELECTRICAL PROPERTIES

	Unit	
μ_0 (1 mc)		50
E (1 mc)		13
*Bs	gauss	1600
*Br	gauss	1300
*Hc	oersted	2.0
Curie Point	$^{\circ}C$	215
Vol. Resistivity	ohm-cm	10^9

* DC Values with Hmax equal 25 oersteds

MICROWAVE PROPERTIES
at X-band

Sample: Dia. 0.250"
Leng. 1.114"
Mounted in waveguide diameter 0.9375"

Absorption Loss	0.3 db
**Saturation Rotation	210°/inch of length

** Varies as a function of frequency, core geometry, and Holding device

ALL DATA BASED ON NOMINAL VALUES

TABLE II

FERRITE SAMPLES

Sample No.	Shape	Diam.	Length	Ferramic
1.	rod	0.378"	3.000"	MF-1331
2.	rod	0.378"	1.250"	MF-1331
3.	rod	0.378"	1.750"	MF-1331
4.	rod	0.230"	2.544"	MF-1331
5.	rod	0.182"	2.979"	MF-1331

APPENDIX II

APPROXIMATE POWER REQUIRED TO DRIVE A SINGLE SIDEBAND MODULATOR

Copper Losses:

Assume magnet field coil has a resistance of 47 ohms.

Assume field current of 0.15 amperes.

$P_c = (0.15)^2 \times 47 = 1.058$ watts

Magnetic core materials:

(1) Permalloy: $H_c = 0.05$ oersteds

$\rho = 16 \times 10^{-6}$ ohm-cm

gage = 0.014 inches

(2) Supermalloy: $H_c = 0.002$ oersteds

$\rho = 60 \times 10^{-6}$ ohm-cm

gage = 0.002 inches

(3) Ferramic MF-1331: $H_c = 2.0$ oersteds

$H_s = 1600$ gauss

$H_r = 1300$ gauss

Hysteresis losses:

Losses for iron cores are computed assuming $B_{max} = 10{,}000$ gauss or saturation induction if less than 10,000 gauss. Actual losses will be less than the computed due to smaller maximum flux densities actually involved. Losses for the ferrite are computed on the basis of a square hysteresis loop. A frequency of 500 cycles per second is assumed. The volume of the iron core is about 500 cubic centimeters and that of the ferrite is about 3.23 cubic centimeters.

Permalloy:

$w = 200$ ergs/cm^3/cycle

$P_h = wfV = 200 \times 500 \times 500 = 50 \times 10^6$ ergs/sec

$P_h = 5$ watts

Supermalloy:

$w = 10$ ergs/cm^3/cycle

$P_h = 10 \times 500 \times 500 = 2.5 \times 10^6$ ergs/sec $= 0.25$ watts

Ferramic MF-1331:

$$w = \frac{4\ B_r\ H_c}{4}$$

$$w = \frac{1300 \times 2.0}{\pi} = 828 \text{ ergs/cm}^3\text{/cycle}$$

$P_{hf} = 828 \times 3.23 \times 500 = 133.8 \times 10^4$ ergs/sec $= 0.1338$ watts

Eddy current losses:

ρ in ohm-meters

a thickness of sheet in meters

B_m maximum flux density (assume 0.5 webers/m^2)

Permalloy:

$$P_e = \left(\frac{\pi^2}{8 \times 16 \times 10^{-4}}\right)\left(0.014 \times 2.54 \times 10^{-2} \times 500 \times 0.5\right)^2 = 6.07 \text{ WATTS/m}^3$$

$P_{et} = 6.07 \times 500 \times 10^{-6} = 3.035 \times 10^{-3}$ watts

Supermalloy:

$$P_e = \left(\frac{\pi^2}{8 \times 60 \times 10^{-4}}\right)\left(0.002 \times 2.54 \times 10^{-2} \times 500 \times 0.5\right)^2 = 3.165 \text{ WATTS/m}^3$$

$P_{et} = 3.165 \times 500 \times 10^{-6} = 1.582 \times 10^{-3}$ watts

Eddy current losses in ferrite are negligible due to extremely high volume resistivity.

Total driving power required:

Permalloy core:

$P_t = 1.058 + 5.00 + 0.003 + 0.134 = 6.195$ watts

Supermalloy core:

$$P_t = 1.058 + 0.250 + 0.002 + 0.134 = 1.444 \text{ watts}$$

It should be emphasized that the above powers required give only order of magnitude results. It is expected that actual losses will vary due to different maximum flux densities in the iron core, due to imperfect insulation between laminations of the iron core, variations in thickness of the laminations and variations in the required magnitude of the magnetizing current.

Sheldon
A broadband microwave single sideband modulator using ferrites ...

28470
Thesis
S442
Sheldon
A broadband microwave single sideband modulator using ferrites ...